Wax

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Packet Radio Temporary Note 33

PROCESSING SPREAD-SPECTRUM SIGNALS

1.0 INTRODUCTION

Much of the past development of spread spectrum systems has been dedicated to providing an anti-jam margin against unlike signals or from intelligent jammers. However, for the packet radio system it appears that the primary merit in considering spread spectrum lies in its potential to provide time discrimination for overlapping signals having the same code. Overlapping signals can originate as the result of time conflicts between two or more users or from multipath.

It is also possible to reduce self-interference or crosstalk between various packet radio users by assigning one group of users different packet data codes from the coding assigned to other group(s). This form of coding provides code discrimination. Each of the code groups would have good auto-correlation properties but the cross-correlation properties of the codes would minimize the affect of one code on the correlator circuits designed for another code. There are a limited number of codes exhibiting these cross-correlation properties, however, unless large spread factors are used.[1][2]

The purpose of this note is to introduce a conceptual spread spectrum receiver model. The model is introduced in order to focus on the some of the radio requirements, if spread spectrum modulation is employed for packet radio. At the same time, it is hoped that this model will provide the network analyists with a concept of the operation of the equipment. It is intended that the model be used as a point of reference for developing a network transmission strategy. The model will also be used to develop cost factors which will be provided in a future note.

The model uses Surface Acoustic Wave Devices (SAWD) for demodulating packet data. These devices were chosen because the matched filter signal processing is done in parallel rather than in series. This simplifies otherwise serious synchronization requirements for packet radio. Use of the SAWD also results in a spread spectrum modem which is economical and simple in operation when moderate spread factors (16 to 127) are employed.

The issue of preamble length is also introduced. The affect of preamble length on the ability of the receiver to acquire synchronization in a gaussian noise environment is considered. It was concluded that a lower bound preamble length equal of 6 bit times is required for reliable synchronization.

A brief discussion of Surface Acoustic Wave Devices is given. Some of the device characteristics which will impact the design of the radio equipment is given. A summary of spread spectrum definitions is given in paragraph 7.

2.0 BASIC SPREAD SPECTRUM RECEIVER MODEL

Data packets consist of a coded SYNC preamble plus up to 1000 coded data bits. The spread spectrum waveform is a PSK type waveform where a pseudo-random code of length N is used to represent a binary one and its one's complement, a binary zero. For synchronization reliability, the preamble length is normally several bit intervals in length. Information is encoded by phase shifting a carrier.

Figure 2-1 illustrates operation of the receiver. An IF carrier (typically 60-300 MHz) is applied to two Surface Acoustic Wave Devices (SAWD) which serve as matched filters. One filter is matched to the preamble code and its output is used for synchronizing the receiver time base to the packet data. The other filter is

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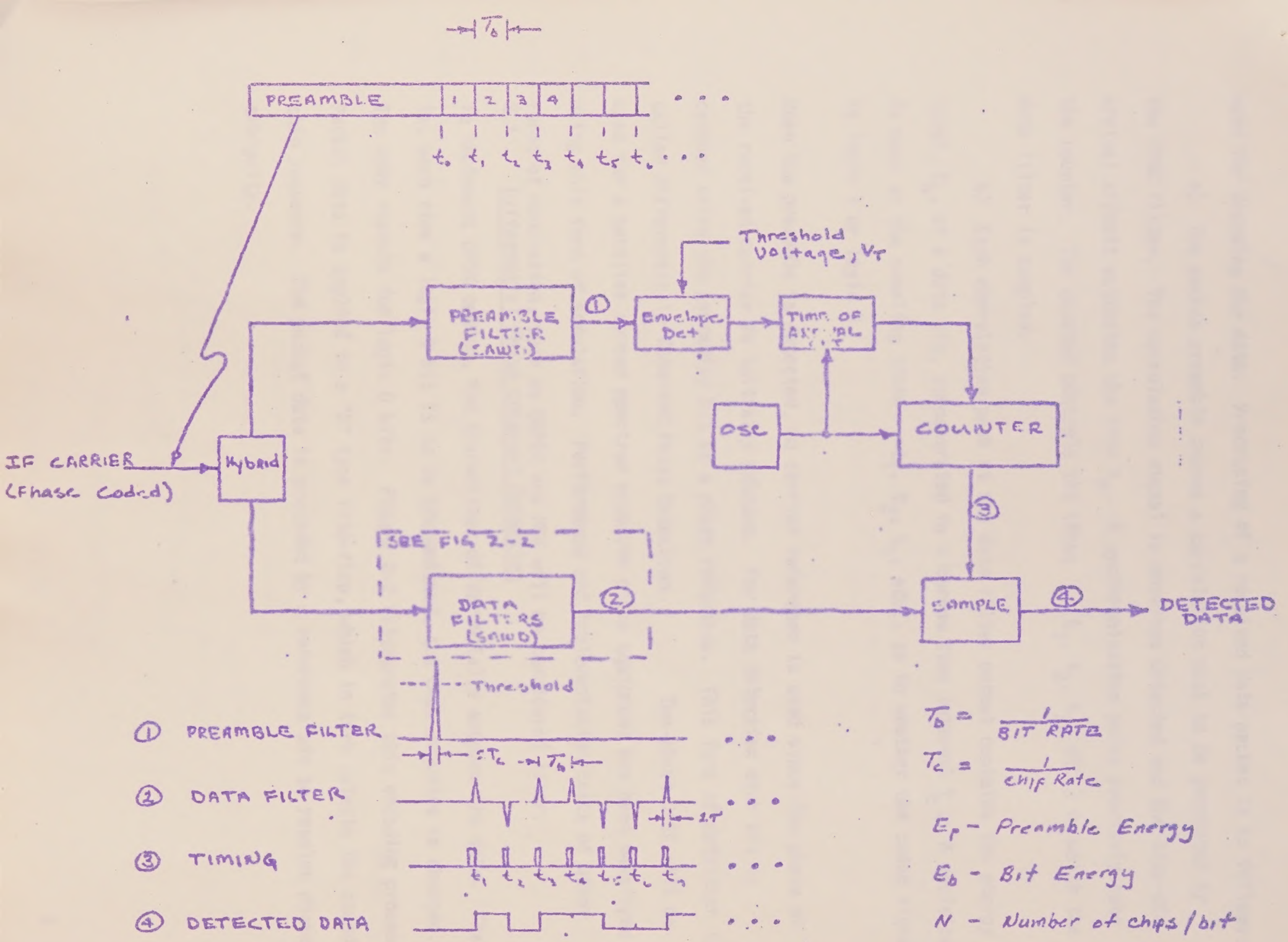
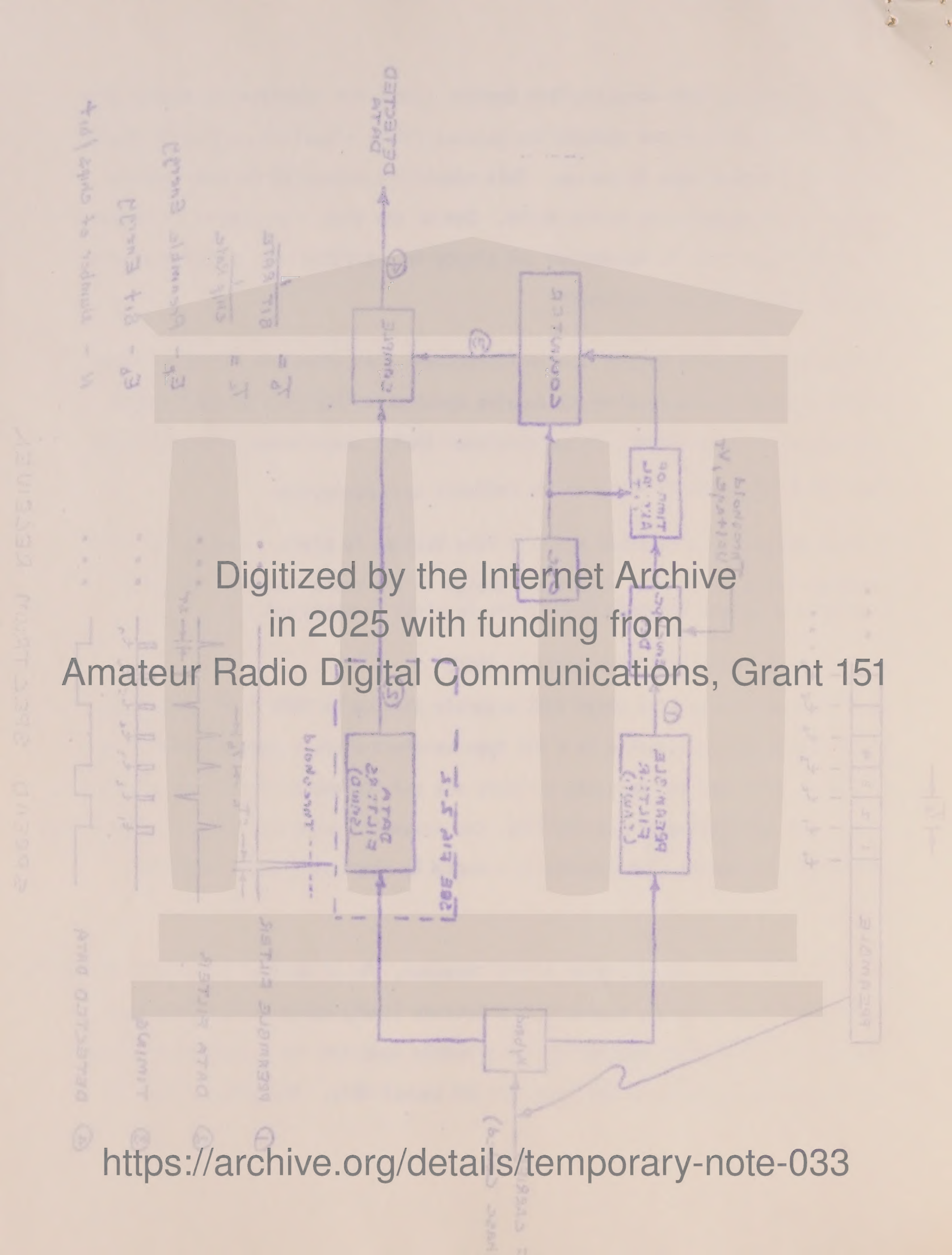


SPREAD SPECTRUM RECEIVER

FIGURE 2-1



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used for decoding the data. Processing of a received data packet is as follows:

a) The packet preamble causes a correlation peak to be generated by the SYNC filter. The correlation signal is envelope detected and the time-of-arrival circuit estimates the time t_0. A synchronization pulse precisely sets the counter. The counter controls the times (t_1, t_2, t_3, etc.) in which the data filter is sampled.

b) Each correlation peak at the data filter output contains the energy level, E_b, of a data bit, concentrated in a narrow time interval T_c. A decision is made at the sampling times t_1, t_2, t_3, etc., as to whether the coded signal is logic 1 or logic 0.

When the preamble is detected, no carrier reference is used since the phase of the received carrier is initially unknown. For data detection each bit is decoded using the preceding bit as a phase reference. This form of detection is called Differentially Coherent Phase Detection. Two-phase DPSK will be used for a baseline spread spectrum receiver since equipment has been developed using this form of modulation. Performance and implementation costs of other forms of modulation, such as QDPSK and MSK will be considered later.

2.1 Differential Phase Coherent Detection

To implement DPSK coding, the transmitter differentially encodes the data. That is, each time a logic 1 bit is to be transmitted, the code sequence is changed. The code repeats for logic 0 bits. Figure 2-2 illustrates this encoding process. Packet data is applied to a "D" type flip-flop, which in turn selects the coded chip sequence. The packet data is preceded by a reference bit to resolve phase ambiguity.

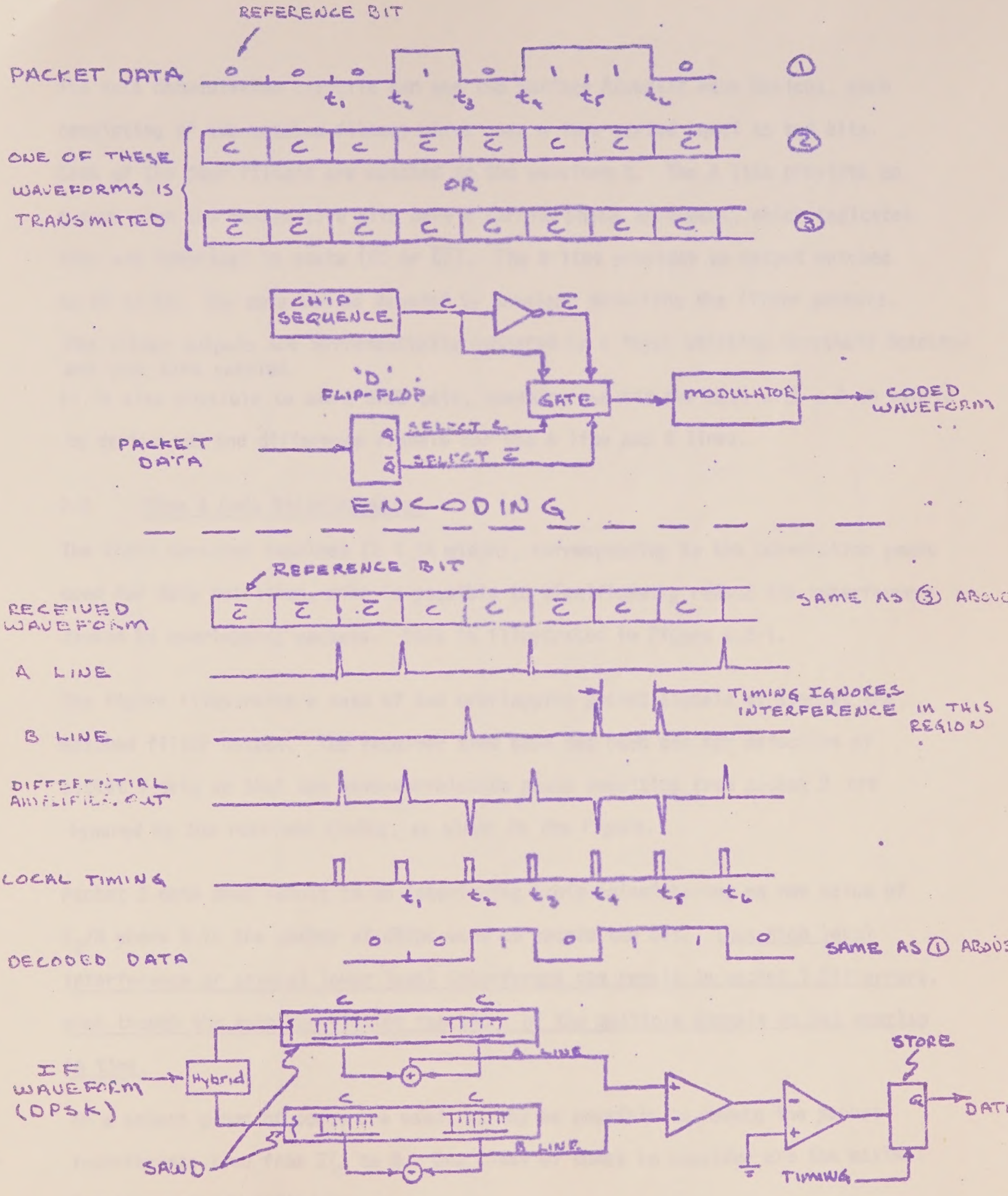
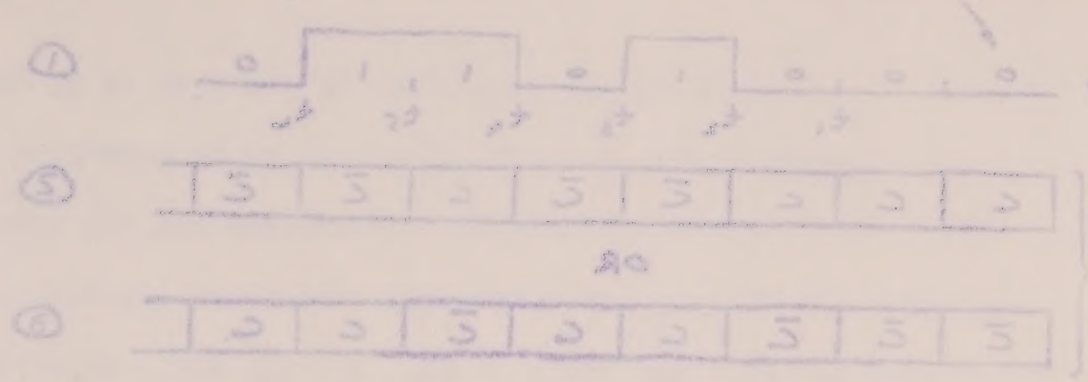
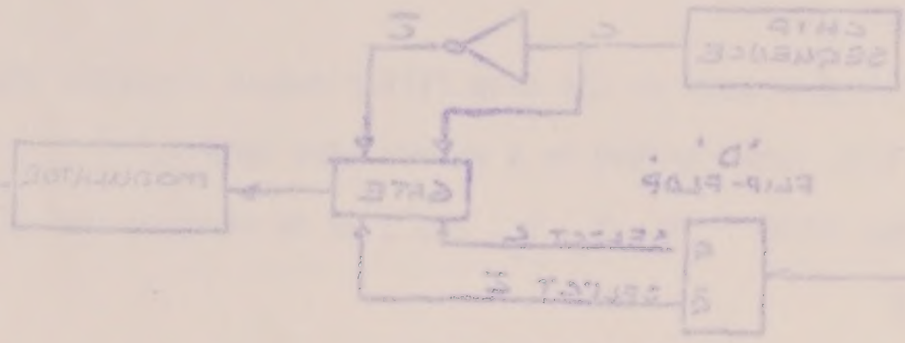
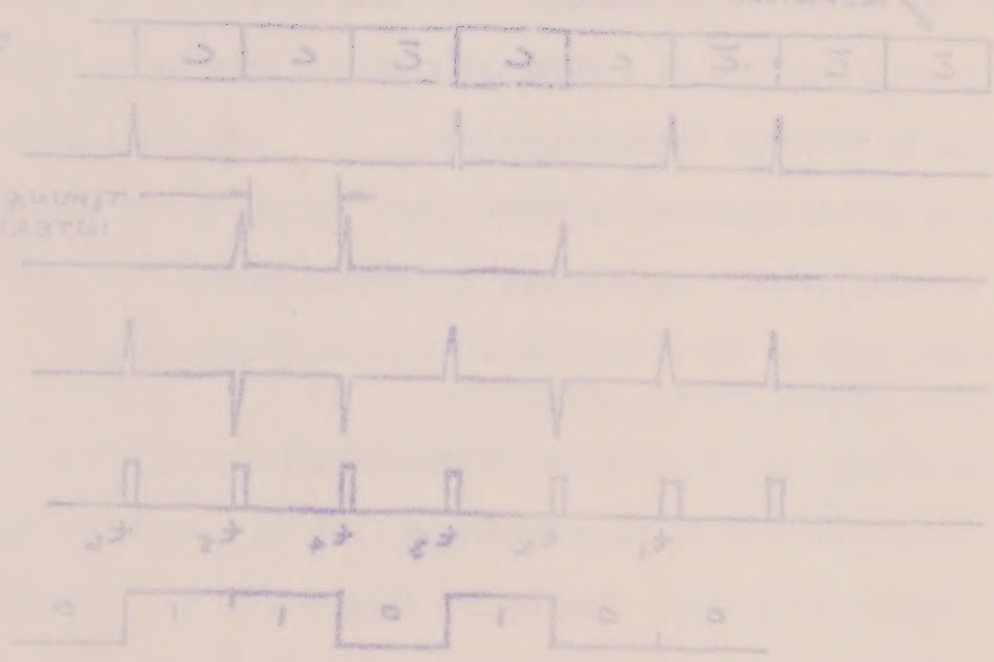
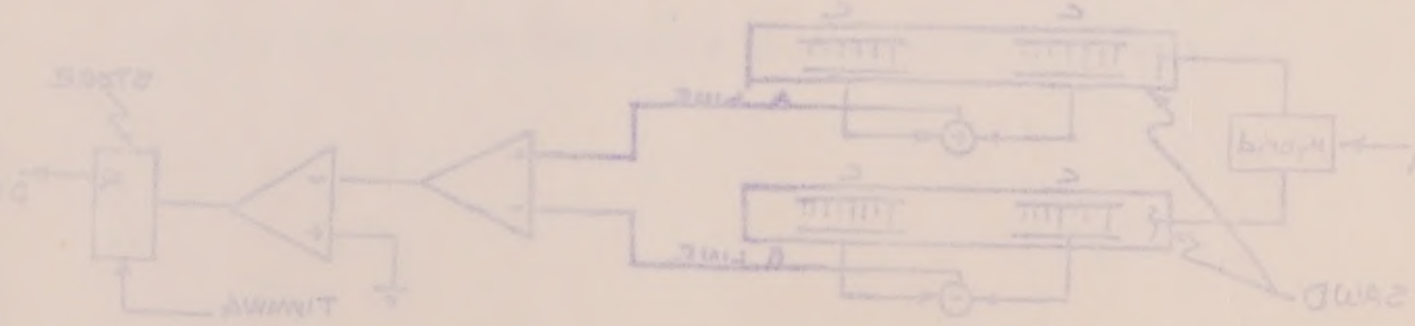

REFERENCE BIT
PACKET DATA
0 0 0 1 0 1 1 0
t_1 t_2 t_3 t_4 t_5 t_6
①
ONE OF THESE WAVEFORMS IS TRANSMITTED
C C C $\bar{C}$ $\bar{C}$ C $\bar{C}$ $\bar{C}$
②
OR
$\bar{C}$ $\bar{C}$ $\bar{C}$ C C $\bar{C}$ C C
③
CHIP SEQUENCE
C
$\bar{C}$
'D' FLIP-FLOP
GATE
MODULATOR
CODED WAVEFORM
PACKET DATA
Q
$\bar{Q}$
SELECT C
SELECT $\bar{C}$
ENCODING
REFERENCE BIT
RECEIVED WAVEFORM
$\bar{C}$ $\bar{C}$ $\bar{C}$ C C $\bar{C}$ C C
SAME AS ③ ABOVE
A LINE
TIMING IGNORES INTERFERENCE IN THIS REGION
B LINE
DIFFERENTIAL AMPLIFIER OUT
LOCAL TIMING
t_1 t_2 t_3 t_4 t_5 t_6
DECODED DATA
0 0 1 0 1 1 0
SAME AS ① ABOVE
C C
IF WAVEFORM (DPSK)
Hybrid
A LINE
B LINE
SAND
STORE
Q
DATA
TIMING
DEMODULATION

FIGURE 2-2









The data demodulation circuits can use two Surface Acoustic Wave Devices, each consisting of two matched filters which span a time period equal to two bits. Each of the four filters are matched to the waveform C. The A line provides an output when two consecutive bits have a carrier phase agreement, which indicates they are identical in state (CC or $\overline{CC}$). The B line provides an output matched to $C\overline{C}$ or $\overline{C}C$. The data can be decoded by envelope detecting the filter outputs. The filter outputs are differentially compared by a level shifting threshold detector and then time sampled.
It is also possible to use a SAWD pair, spanning two bit periods, with a 3 db hybrid to derive sum and difference signals for the A line and B lines.

2.2 Time & Code Discrimination

The short duration impulses (2 T_c in width), corresponding to the correlation peaks used for data detection, make it possible to significantly reduce the interference caused by overlapping packets. This is illustrated in Figure 2.2-1.

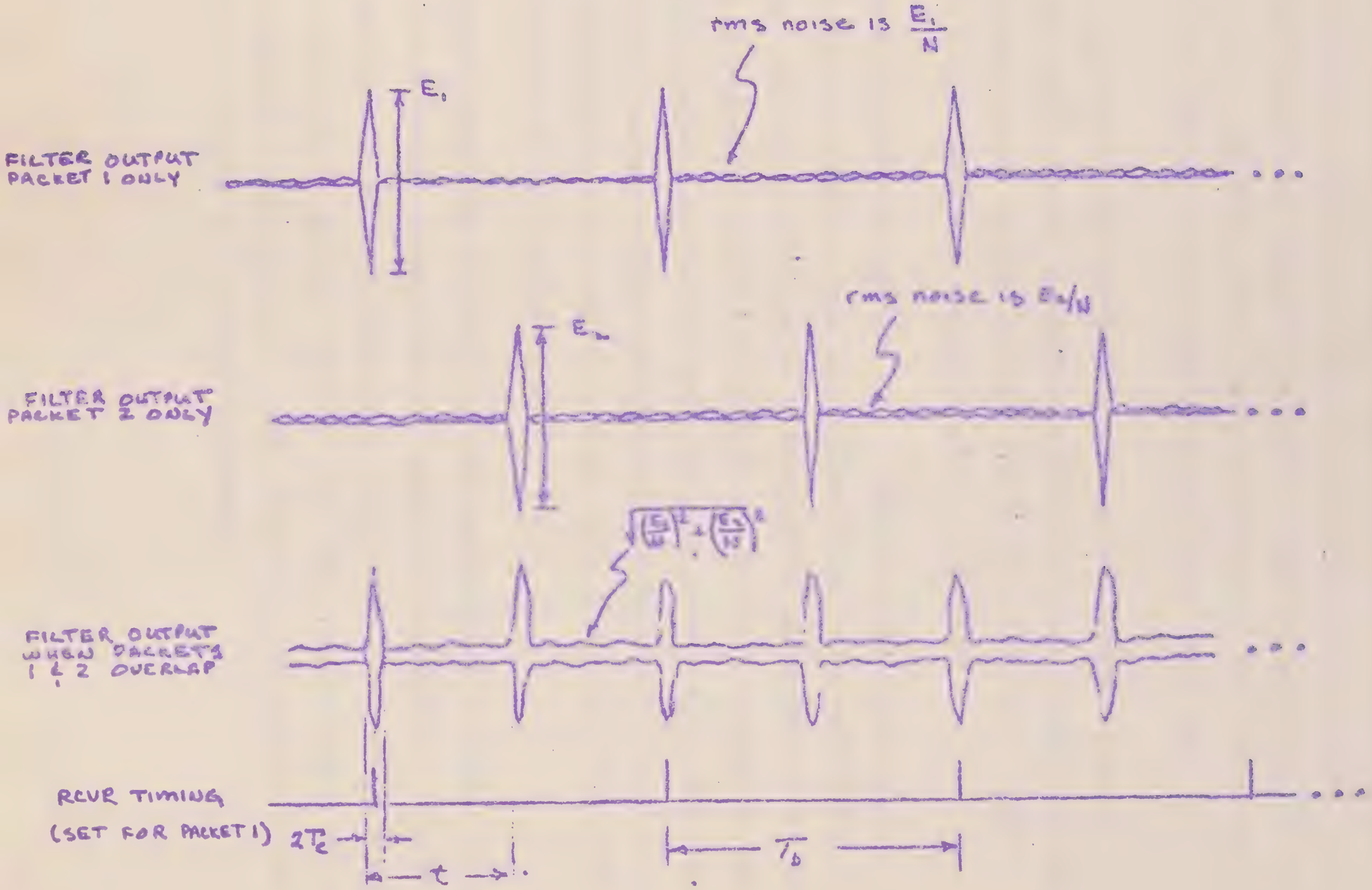
The figure illustrates a case of two overlapping packet signals as seen at the matched filter output. The receiver time base has been set for detection of packet 1 data so that the auto-correlation peaks resulting from packet 2 are ignored by the receiver timing, as shown in the figure.

Packet 2 data does result in an interfering "chip noise" having an rms value of E_c/N where N is the number of chips used to encode the bit. Thus high level interference or several lower level interferors can result in packet 1 bit errors, even though the auto-correlation functions of the multiple signals do not overlap in time.

If a select group of codes are used, it may be possible to reduce the packet interference time from $2T_c$ to 0. One class of codes to consider are the maximal length codes identified by:

$$\frac{\emptyset\,(2^k - 1)}{k}$$

where: $\emptyset(x)$ is the Euler phi-function.



TIME DISCRIMINATION

FIGURE 2.2-1

The cross-correlation function for two of these codes, 31 chips in length, is shown in Figure 2.2-2. The figure illustrates the caution which must be used when attempting to apply code discrimination in a random access network. The RMS interference level between two overlapping packets has been increased to $6.5\left(\frac{E_2}{\sqrt{N}}\right)$ vs. (E_2/N) for packets having the same code. The relative cross-correlation interference can be reduced by using longer length codes.

Spread spectrum modulation should also make it possible to reject specular multipath signals arriving at the receiver later than 2 chip intervals from the start of the message. Specular multipath will result in interfering signals similar to those produced by overlapping packet transmissions.

In summary it is felt that the time discrimination provided by spread spectrum can significantly reduce the self-interference in the packet radio network. Transmission strategies can be developed for random packets having the same code which enable detection of the first arriving packet provided the start of a second packet is not received during the short interval, $2T_c$. It may also be reasonable to consider equipping the repeater with multiple gating functions so that overlapping packets, having the same code, could be processed concurrently when favorable multipath conditions exist.

It should be pointed out that the use of large spread factors for minimizing the interference time, $2T_c$ increases receiver complexity and bandwidth utilization. For this reason the cost and performance of a baseline receiver having a spread factor of 15 will be considered initially. Cost and performance charts will then be developed for spread factors up to 1000. The merits of code discrimination will also be analyzed in a more definitive manner.

[illegible] Figure 2 [illegible]. The figure illustrates the caution which must be used [illegible] to apply code discrimination [illegible]. The interference level between two overlapping packets has been increased to [illegible] vs. (E_2/N) for packets having the same code. The [illegible] interference can be reduced by using [illegible] length codes.

[illegible] modulation should also make it possible to [illegible] signals arriving at the receiver later than 2 chip [illegible] message. Specular multipath will result in [illegible] produced by overlapping packet transmissions.

[illegible] that the time [illegible] discrimination [illegible] significantly reduce the self-interference in the [illegible] strategies can be developed for random [illegible] enable detection of the first arriving packet [illegible] not received during the short interval, 2 [illegible] consider equipping the repeater with multiple [illegible] packets, having the same code, could be produced [illegible] multipath conditions [illegible].

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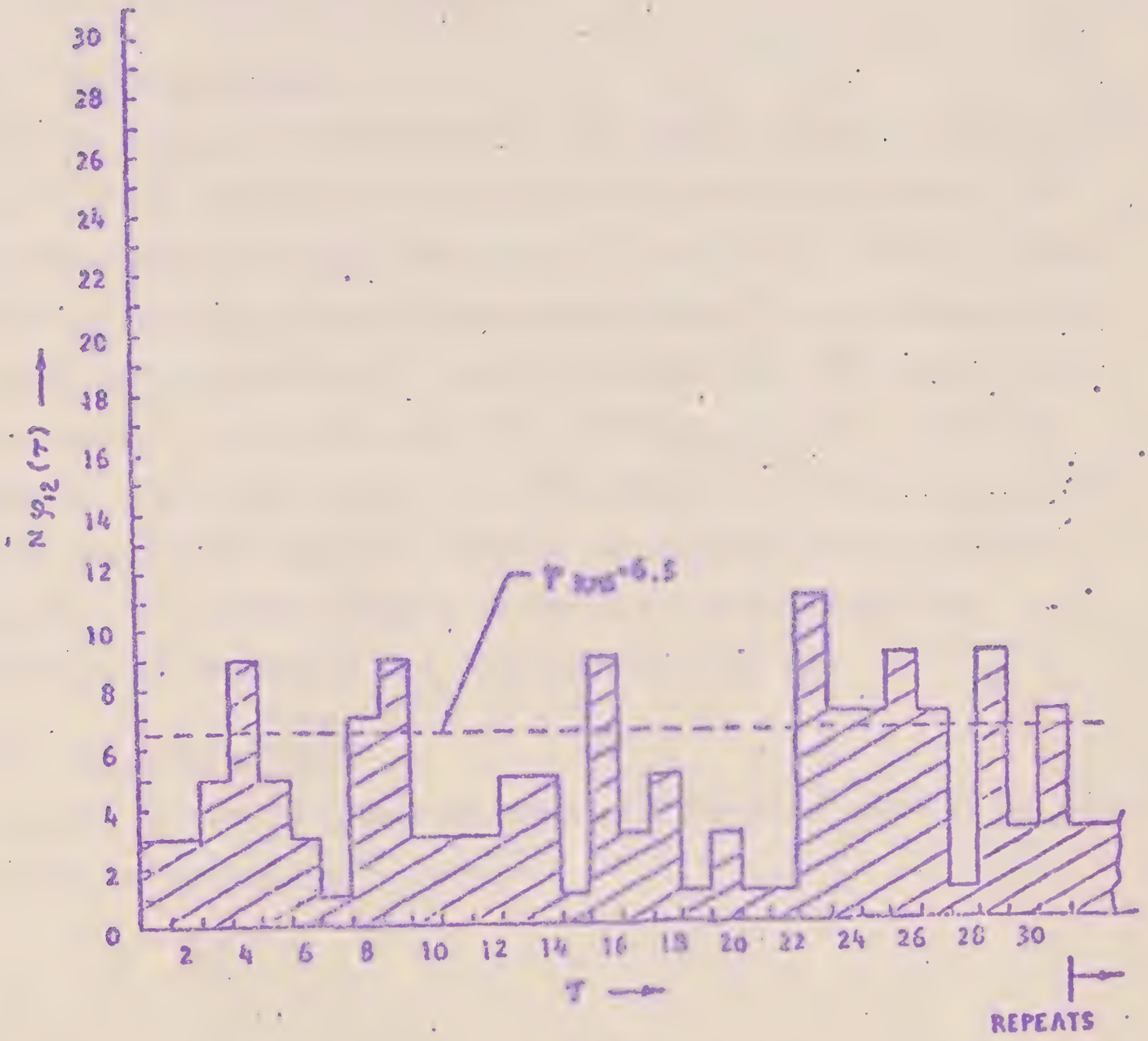
CODE 1: 11111 00011 01110 10100 00100 101100

CODE 2: 11111 00110 10010 00010 10111 011000



Cross-Correlation Function Between Codes No. 1 and No. 2

FIGURE 2.2-2

3.0 PREAMBLE LENGTH

The criteria for determining the necessary preamble length for data packets are:

1) The noise environment in which sync must be detected.
2) The probability established for preamble detection.
3) The acceptable false alarm rate established for the sync detection circuits.
4) The accuracy with which the packet time of arrival must be defined for reliable data detection.

3.1 Noise Environment

For a 1000 bit packet to be correctly received, 99% of the time, a bit error rate of 10^{-5} is required if no errors are to be allowed in the packet. For DPSK, this establishes a data threshold for $\frac{E_b}{N_o}$ at 10.5 db. However, it would be well to consider providing reliable synchronization for environments having a lower signal-to-noise ratio. The 1000 bit packet has a 50% probability of being correctly received for a BER = 7×10^{-4} (E_b/N_o = 8.2 db). A 200 bit packet has a 50% chance for BER = 4×10^{-3}, which corresponds to E_b/N_o = 6.5 db. If the same preamble length was specified for both packet sizes, it would be desirable from a system standpoint to provide a preamble having a 99% probability of detection in an E_b/N_o environment of 6.5 db.

3.2 False Alarm Time[2]

If an envelope detector is used for detection of the packet preamble, the probability of false detection of sync is given by

$$P_{fa} = \frac{1}{T_{fa}B} \qquad (1)$$

(1)

where

P_{fa} is the probability of acquiring false sync

T_{fa} is the average time between false syncs

B is the information bandwidth when spread spectrum processing is considered.

The mean time between false synchronization is related to the threshold setting of the envelope detector and the receiver noise level.

$$T_{fa} = \frac{1}{B} \exp \frac{V_T^2}{2N_o} \qquad (2)$$

where

V_T is the threshold voltage

N_o is the mean-square noise voltage in the information bandwidth

When false synchronization occurs, the receiver (capable of receiving one packet at a time) is blocked for one packet interval. How often should we allow this to occur? Equation 1 is calculated for various number of packet intervals below.

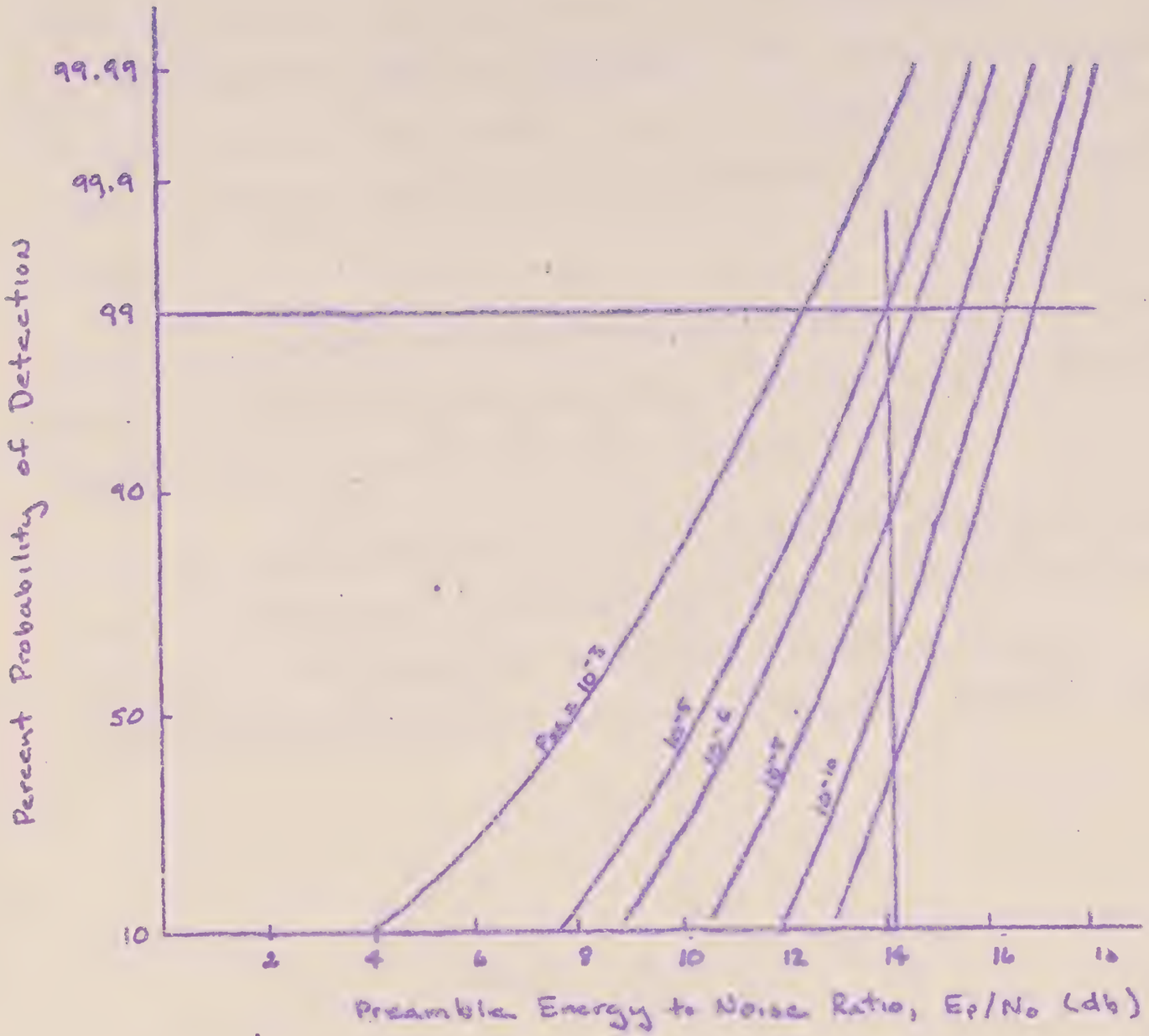
Number of 1000 Bit Packet Intervals	T_{fa}	$P_{fa} = \frac{1}{T_{fa}(200 \times 10^3)}$
50	.5 sec	10^{-4}
500	5 sec	10^{-5}
5000	50 sec	10^{-6}
50,000	500 sec	10^{-7}

Since a receiver can only go into a false sync mode during idle periods, negligible impact of system throughput would result for P_{fa} as high as 10^{-5}. I will use $P_{fa} = 10^{-5}$ as an upper bound for false sync probability.

T_{fa}	$P_{fa} = T_{fa}$
.5 sec	10^{-4}
5 sec	10^{-5}
50 sec	10^{-6}
500 sec	10^{-7}

3.3 Probability of Sync Detection

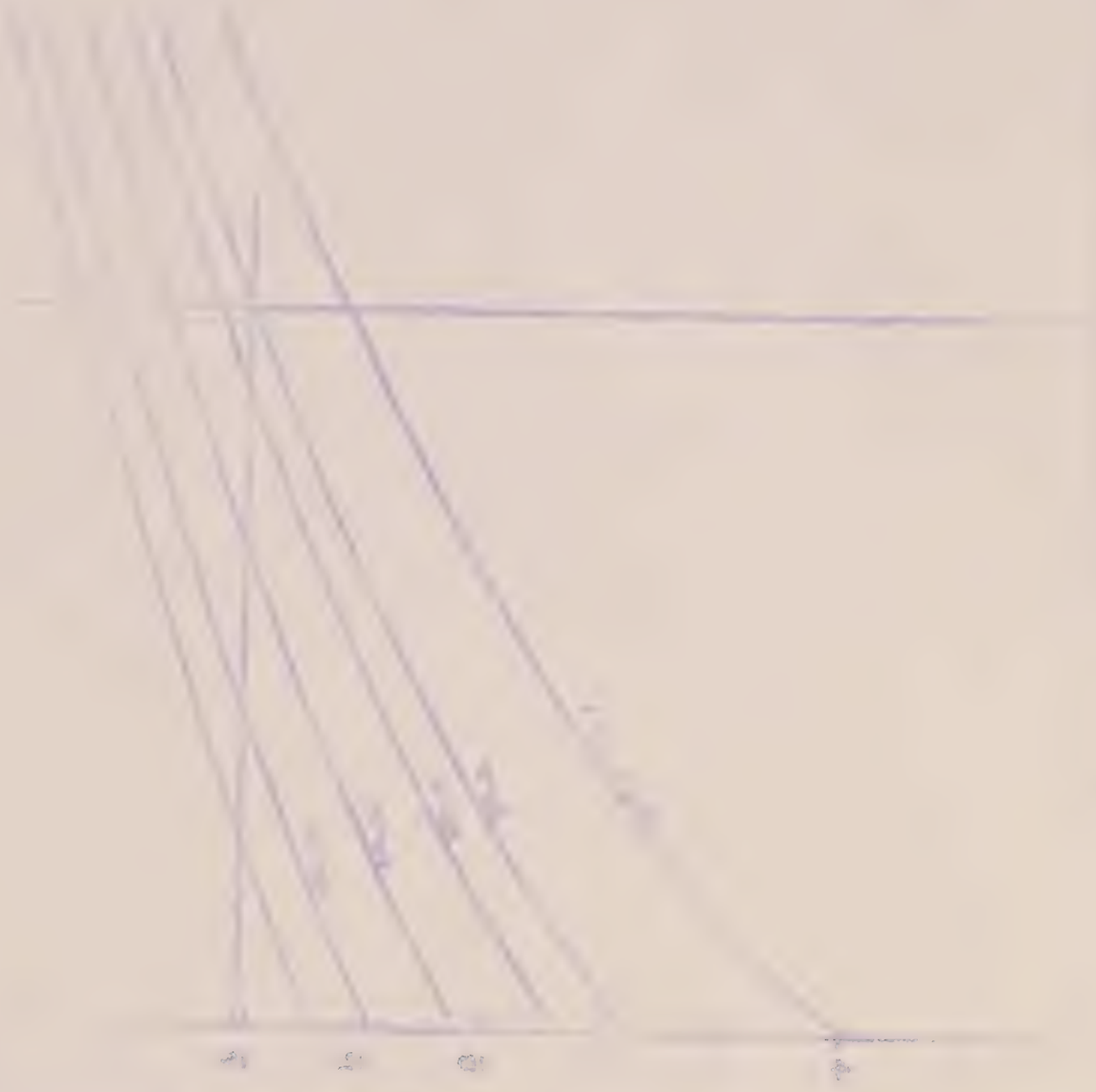
Figure 3-1 is a family of curves relating probability of preamble detection to the preamble energy and Gaussian noise levels. (These curves are taken from reference 3, pg. 34.) I have taken as a lower bound a requirement of 99% probability for sync detection and a false alarm probability of 10^{-5}. Under these conditions, a preamble energy-to-noise level of 14.2 db is required. The curves of figure 3-1 provide an indication of the preamble length required for reliable synchronization by a user operating at or near a noise margin providing a BER of 4×10^{-3}. The preamble energy level should be 14.2 - 6 = 7.7 db higher than the energy per bit. This lower bound energy level can be provided by using 6 bit intervals for the synchronization preamble.



Note: Curves show probability for detection of a sine wave in noise. See reference 3

PREAMBLE ENERGY vs PROBABILITY OF DETECTION

FIGURE 3-1



3.4 Use of the Preamble for Defining Packet Time of Arrival

The narrow auto-correlation pulse at the output of the preamble matched filter can be used for adjusting the receiver time base for sampling the packet bits. Ideally data sampling should occur during the narrow time interval, $2T_c$, in order to minimize the effects of interference. This implies a precise setting of receiver timing. If the preamble is used to set timing, the accuracy of the setting is determined by the accuracy of the time-of-arrival estimate made on preamble.

The ability to estimate the time of arrival is theoretically limited by noise, interference (such as multipath), the energy of the SYNC preamble waveform (E_p), the chip interval T_c, and the degree of waveform bandlimiting.

One method for estimating time-of-arrival is to threshold detect the correlation peak from the SYNC matched filter and derive a time of arrival estimate based on the leading and trailing edges of the correlation peak as shown below. This method of estimating is somewhat immune to small changes in the threshold level.

SYNC AUTO-CORRELATION PEAK

Theoretical work by Swerling and Skolnik (see references 4 & 5) provides lower bound estimation errors for time-of-arrival in a noise environment. Practical estimator circuits, however, only approach these lower bound limits. For reference figure 3-2 is given which shows time of arrival variance as a function of preamble energy to noise ratio. The curves show variances for a chip rate of 5 mcps (T_c = 200 ns). Curve 1 is Swerlings lower bound limit which applies for no bandlimiting (square chips). Curve 2 was developed by North American Rockwell for a center of pulse estimator and would apply for a 3 db RF bandwidth of 10 MHz. The graph also notes the preamble energy to noise levels for a 6 bit preamble in an error environment providing BER of 10^{-5} and 4×10^{-3}. . A 1σ variance between 10 and 20 ns can be expected for these two environments. 3 σ variances, however, would be more appropriate to consider when planning a data detection method.

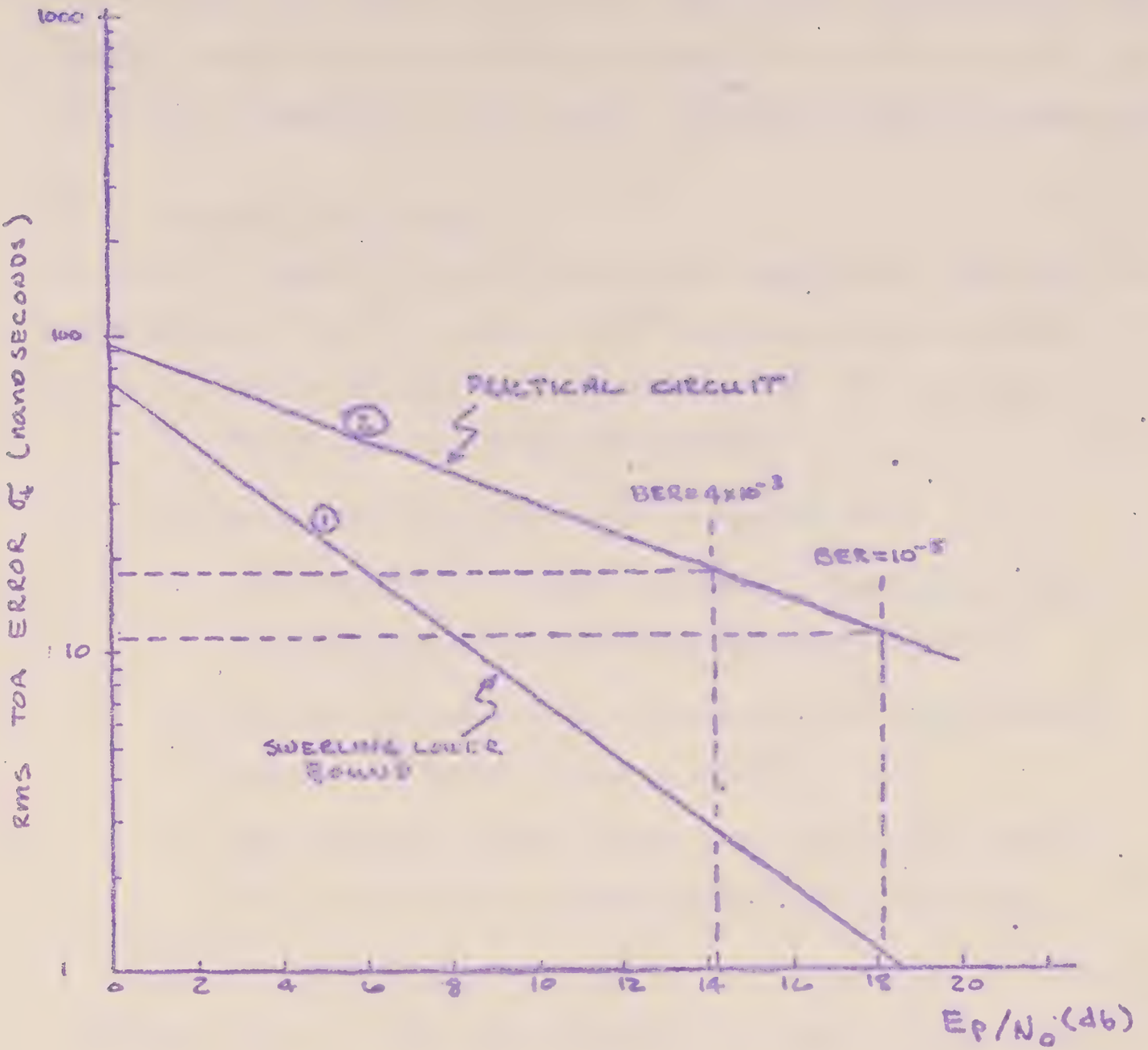
Variances due to clock drift with respect to the transmitted data rate must also be considered, if an asynchronous receiver clock is to be used.

Clock Accuracy	Drift error per 10 msec(1 Packet)
10^{-4}	1000 nanoseconds
10^{-5}	100 nanoseconds
10^{-6}	10 nanoseconds
10^{-7}	1 nanosecond

It can be seen (for the 5 mcps case) that unless accurate clocks are used variances due to clock drift can be a significant fraction of a chip interval.

Several data detection strategies are now under consideration. One method involves the establishment of a timing window which precedes the expected time of arrival of the data auto-correlation pulses. The window is opened approximately one T_c interval before a bit decision is to be made. The outputs from the data

Accuracy	Drift error per 10 min
10^{-4}	1000 nanoseconds
10^{-5}	100 nanoseconds
10^{-6}	10 nanoseconds
10^{-7}	1 nanosecond



Chip rate = 5 mcps,

E_p - Preamble energy

N_o - rms noise

① $\sigma^2 = \frac{T^2}{2\left(2\frac{E}{N_o}\right)^2}$ ② $\sigma^2 = \frac{T^2}{2(2E/N_o)}$

TIME OF ARRIVAL VARIANCE

FIGURE 3-2

delay lines are envelope detected and a threshold detector provides an indication that a bit has been received. The bit is stored and the window is closed until receiver timing indicates it is time for another bit decision to be made. For this method of detection, a 6 bit preamble will provide sufficient timing accuracy.

3.5 Preamble Length Summary

The preamble length was reviewed in terms of the expected noise environment for 50% throughput of a 200 bit packet. The following assumptions were made.

1. The same preamble length would be used for 200 bit packets and 1000 bit packets. Data rate is 100 kbs.
2. Performance would be determined by Gaussian noise.
3. A 99% probability of preamble detection is required for error environments yielding a 50% chance for packet throughput.
4. Under the conditions of (3), the average number of packet times between false synchronization is 500 ($P_{fa} = 10^{-5}$).
5. Time of arrival variances of less than 10% of a chip interval will provide sufficient timing accuracy for data detection.
6. Ideal processing gain can be achieved from the preamble filter.

Using these criteria, I have concluded that for the spread spectrum model, a lower bound preamble length of 6 bit intervals is desirable for reliable synchronization.

4.0 CHARACTERISTICS OF SURFACE WAVE TAPPED DELAY LINES.

This section provides a brief description of some of the principles of Surface Acoustic Wave Tapped Delay lines. Some preliminary design parameters associated with these devices are given which illustrate the nature of the engineering design trade-offs to be made when implementing these devices into a spread spectrum system.

4.1 Principles

The principles of the SAW TDL matched filter for phase coded CW waveforms are illustrated in Figure 4.1. An incoming phase coded waveform, in this case a 7-bit Barker phase code, is fed into the input interdigital transducer (IDT) which converts it into a traveling surface acoustic wave in the piezoelectric substrate. The electric field associated with the acoustic wave in piezoelectric materials allows the signal to be sampled (tapped) at multiple points in the delay path by other IDT's (taps). The taps in Figure 4.1 are placed at precise identical time intervals equal to the chip period of the incoming waveform.

The physical spacing between taps depends on the surface wave velocity (v_s) of the substrate materials and the chip rate of the incoming waveform. For a 10 MHz chip rate on temperature stable ST cut quartz ($v_s = 3.157 \times 10^5$ cm/sec) the tap spacing is 12.49 mils with a normal accuracy of placement of ≤ 10 microinches.

The taps are phase coded so that when the waveform is completely loaded into the TDL all tap outputs add in phase, giving the correct correlation function. Figure 4.1 shows an external phase shift network to produce 180 degrees phase shifts prior to summation. Other techniques of tap phase coding are illustrated in Figure 4.2 where 4-2a and 4-2b are fixed coding techniques. Figure 4-2c

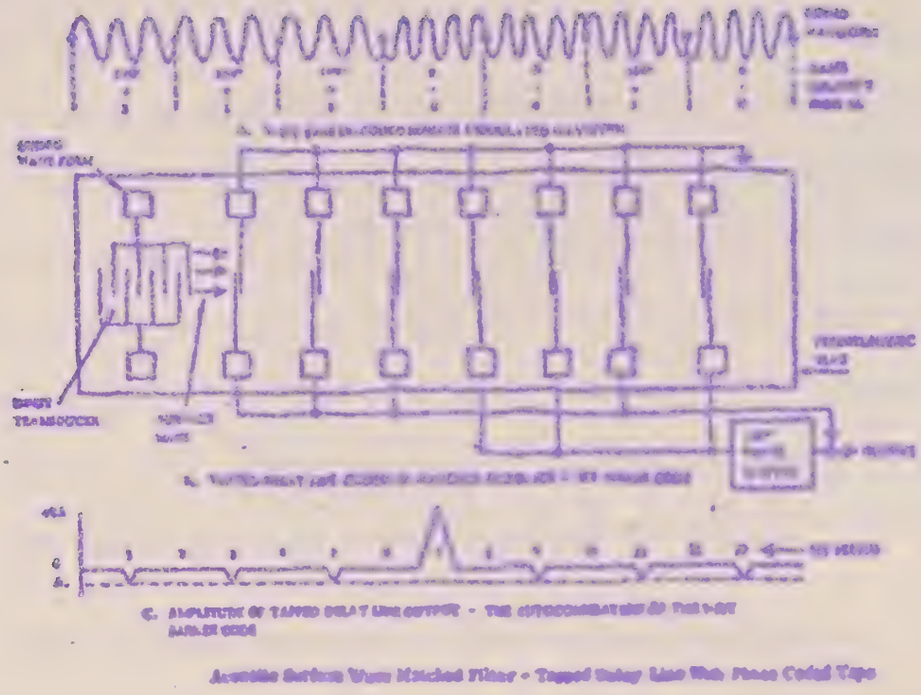


FIGURE 4-1

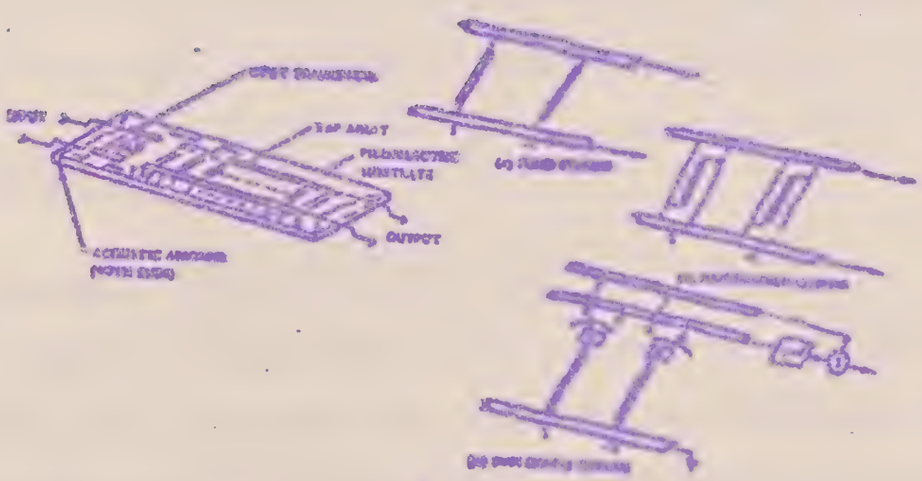
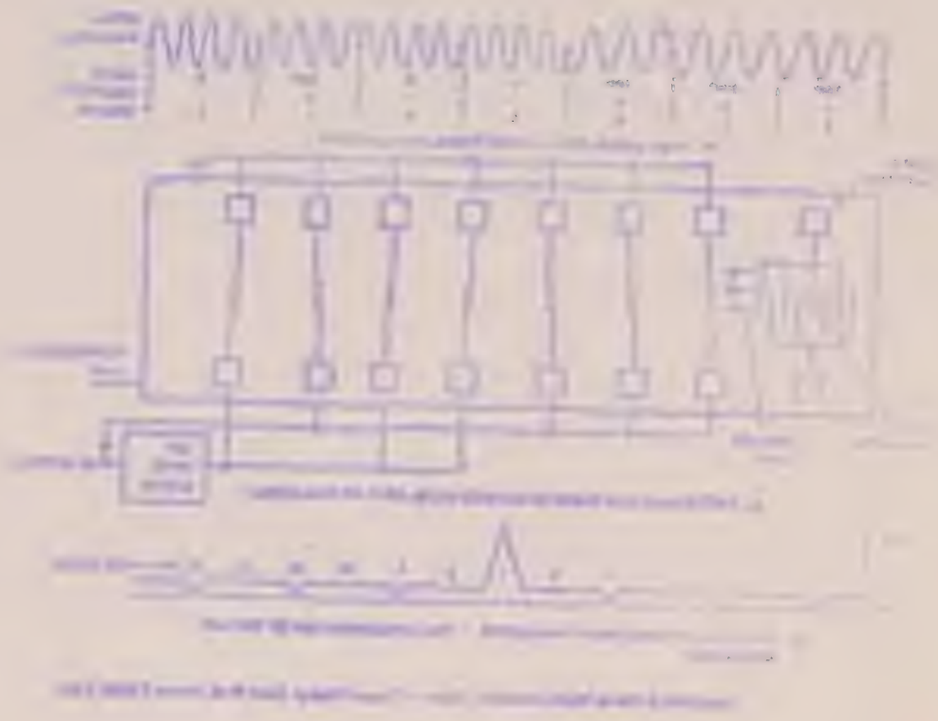
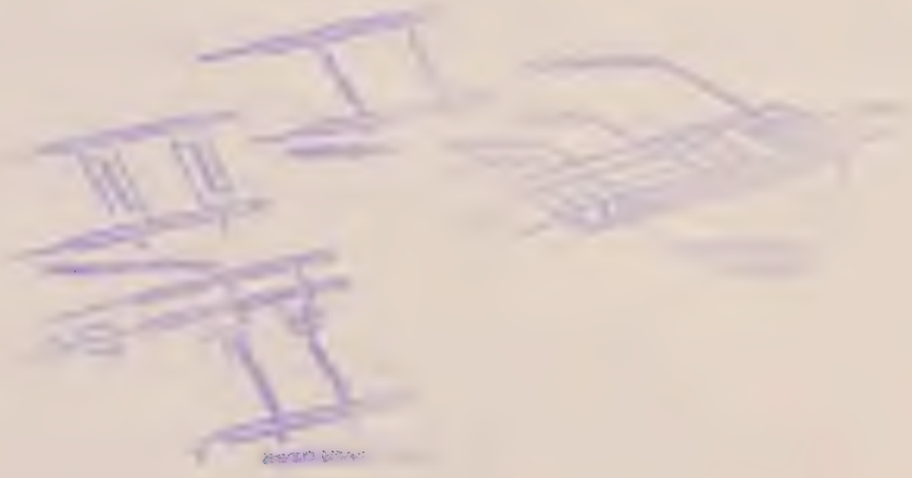


FIGURE 4-2





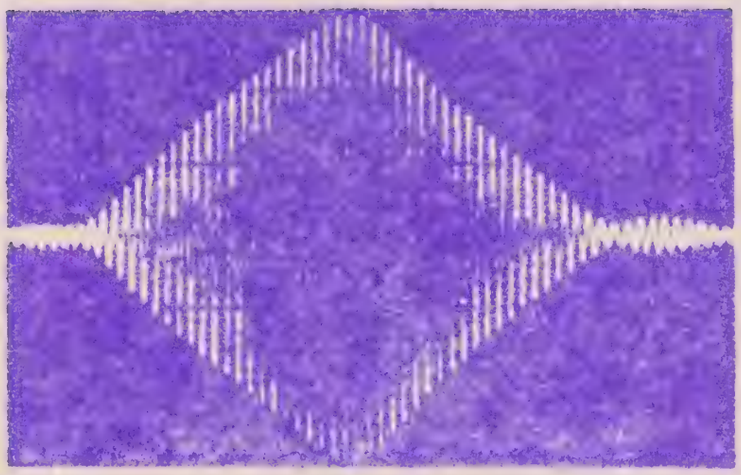
represents a switchable tap approach. For the configuration of Figure 4-2a, the tap coding is built into a photomask used to fabricate the SAW TDL and would normally only be used for large production runs of identical coded lines. Figure 4-2b shows a technique whereby TDL's can be fabricated with both sides of the tap connected to both sum lines and the phase coding achieved by opening two lines per tap - as indicated in the schematic. This approach allows large numbers of devices to be fabricated with identical structure and the individual line coding to be carried out quickly and conveniently by techniques such as laser beam burn out on a tape progrewmed XY table.

The transfer function of the surface wave device will exhibit the response required to be a matched filter for each chip of a square modulated waveform. The response of the taps is relatively wideband and so does not distort the waveform. Figure 3.3 illustrates the main peak of the auto-correlation function of a 63 chip, 5 mcps tapped delay line operating at 120 MHz. To first order, this function is the ideal triangular shape expected for a square chip, although some slight rounding is observed at the peak. It was estimated that this amount of round-off would be incurred if the sequence were passed through a single pole filter with 3 dB BW equal to about 30 MHz.

5.2 Device Characteristics

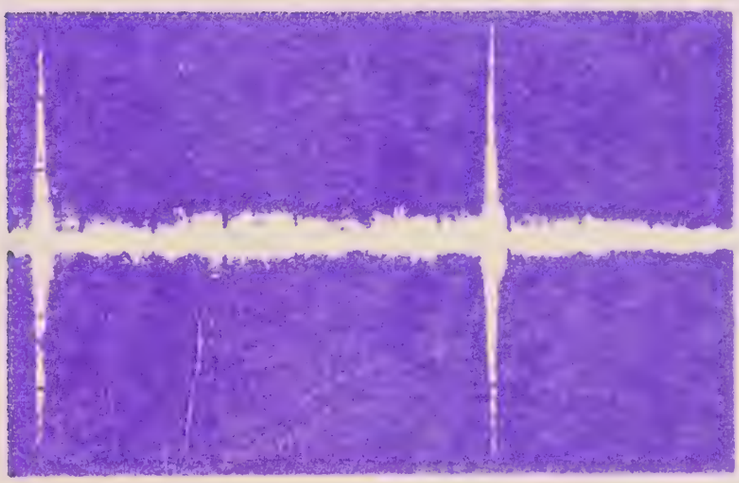
5.2.1 Processing Gain and Insertion Loss

In designing tapped delay lines, one key trade-off is between regeneration and insertion loss. Regeneration occurs when the signal appearing at the load is sufficiently large to cause secondary signals to flow in the delay line. If the impedance of the load is comparable to the impedance of the array of taps, a significant portion of the available power will be reintroduced into the line



Vertical: 0.2 Volts/Division
Horizontal: 50 Nanoseconds/Division

Tapped Delay Line Matched Filter Output



Horizontal: 2 Microseconds/Division

Periodic Autocorrelation Function 63 Chip Sequence

FIGURE 3-3

instead of delivered into the load. Normal practice to reduce this problem is to design the taps so that receive portion of the tap impedance is at least 5 x the resistive portion. This reduces the regenerated signal to a level about 15 dB or more below the main signal. The insertion loss of the SAW TDL from input waveform to correlation peak is normally 25 dB for "optimum" performance.

Figure 3.3 illustrates the slight degradation that can result from these reflections in a tapped delay line. The function shown in Figure 3.3 is the complete (periodic) autocorrelation function of the 63 chip, 5 mcps line. Ideally, since the sequence selected was maximal length, the peak to sidelobe ratio should be 63:1 or 36 dB. In actuality the ratio is seen to be about 63:4 or 24 dB. This value of 24 dB may be considered the average of all accumulated reflected energy.

5.2.2 Freouency Uncertainty

A major factor which can contribute to loss of processing gain is frequency offset of the coded waveform. Figure 3-4 is a curve illustrating the loss in processing gain vs. frequency error. A null frequency (providing infinite loss) occurs at a frequency offset equal to the reciprocal of the time delay of the matched filter. As an example, a line 64 usec long would lose 3 db in processing gain if the frequency offset were 0.45 x 1/64 usec, or about 7 KHz.

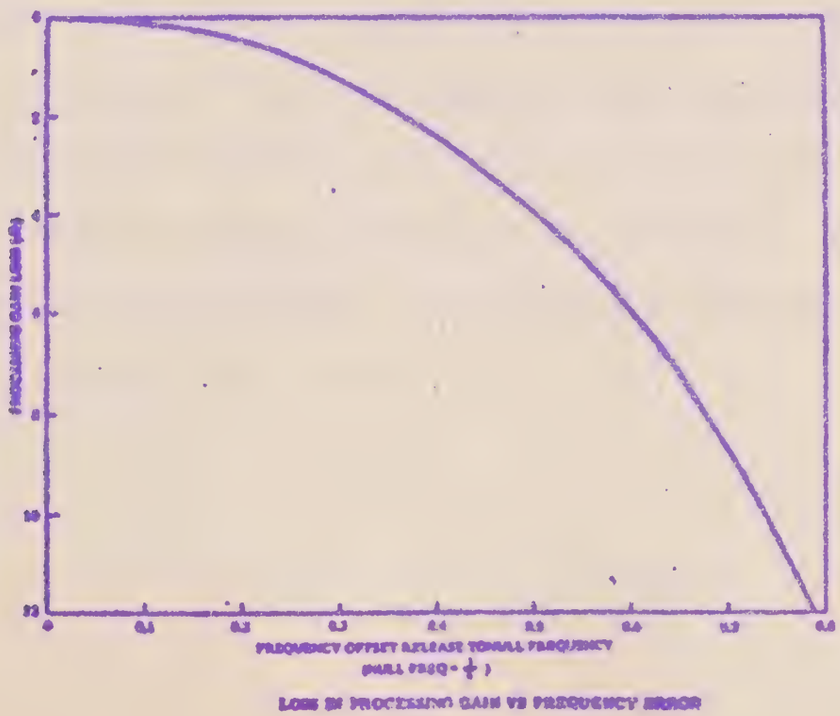


FIGURE 3-4

Three sources of frequency error are the transmitter L.O., receiver L.O. and the matched filter (SAWD) used for sync detection. Doppler must also be considered. Worst case frequency errors for a 1 GHZ radio system with sources stabilized to 10 ppm and .1 ppm are given below.

L.O. SOURCE STABILITY	-50°C to +65°C		0°C to +50°C	
	± 10 ppm	± .1 ppm	± 10 ppm	± .1 ppm
Transmit L.O. (1GHz)	1.0	.1	1.0	.1
Receive L.O. (1GHz)	1.0	.1	1.0	.1
SAWD*	6.0	6.0	2.4	2.4
Doppler (Mach 1)	.44	.44	.44	.44
Worst Case Error	8.44kHz	6.64kHz	4.84kHz	3.04KhZ

*SAWD error based on 50 ppm, fo = 120 MHz for -50°C to +65°C and 20 ppm for 0°C to +50°C.

One means of avoiding the loss effect due to frequency mismatch is to provide several delay lines in series, each operating coherently on a short segment of a sync sequence. Each filter output is detected and then non-coherently summed. As long as the number of lines is small, the loss in overall processing gain is reduced. Thus if four delay lines, each having a line length of 25 μsec were connected in series, a 6 kHz frequency error would result in approximately a 0.2 db loss in processing gain for each line. This would result in an overall loss of processing gain of 0.8 db plus a loss due to the non-coherent addition of each of the lines.

5.2.3 Device Fabrication[6]

Texas Instruments has supplied us with a graph summarizing their fabrication limitations. The graph is attached for your reference . A quote from reference 6, relating to the attached graph is as follows:

"A 25 percent bandwidth has been assumed for practical reasons (insertion loss, spurious responses from device and electronics, etc.). Devices within the center shaded region are readily obtainable in large quantities. Devices in the lightly shaded region are also readily available but require more custom work. The outer shaded region contains those devices which are technically feasible and potentially producible in quantity but still in the development stage. Extrapolations outside this region entail considerable technical as well as economic risk. The top and right hand boundary determine the maximum bandwidth and time lengths which may be obtained using the given level of technology. The product is the maximum chip length obtainable and is shown in the figure for each of the three regions."

The curves developed by TI provide a good indication of the spread factors we should consider for spread spectrum equipment implemented using Surface Wave devices. We should think in terms of 100 chips per bit or less initially. Devices in the 100 chip to 1000 chip area are possible, but are still in the developmental stage.

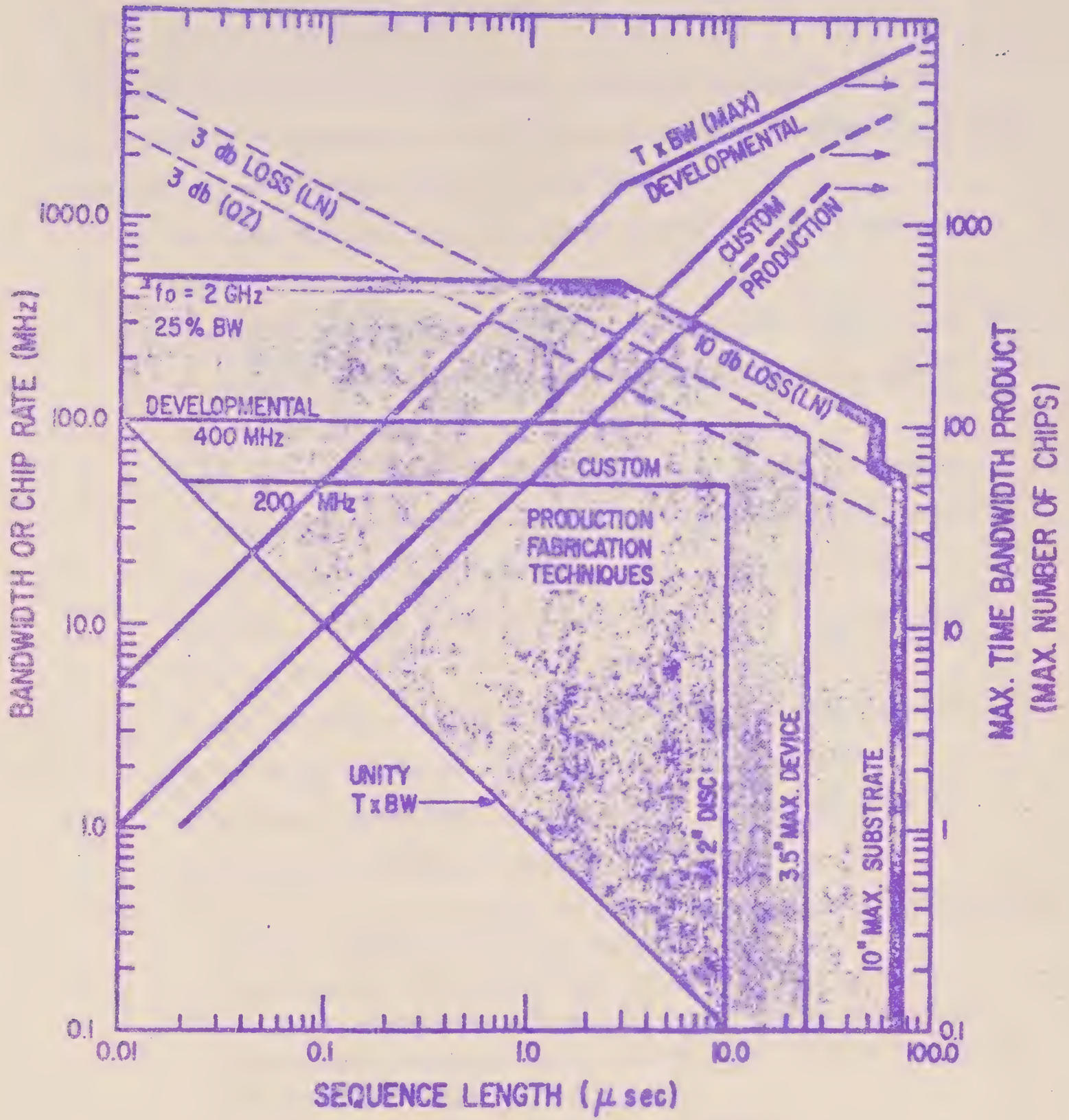


Figure 9. Current Fabrication Capabilities for Coded Devices

7538-4

6.0 SUMMARY OF FUTURE TASKS

The model introduced in this note raises a number of issues which must be considered before performance and cost of spread spectrum equipment can be fully evaluated. Some of these issues are listed below:

1) Manufacturing costs, as a function of chip rate, are required for Surface Acoustic Wave Devices.

2) The sensitivity of processing gain to propagational effects, such as multipath must be measured. This can be facilitated by some propagational tests.

3) A more definitive study of the equipment back-to-back performance is required when various synchronization techniques are considered.

4) Further investigation of cross-correlation coding methods is required. Emphasis should be placed on the use of moderate length codes (16 to 128).

5) A complete receiver model, which includes RF, is needed. Requirements for AGC, bandwidth, linear dynamic range, etc. must be established. Cost curves for the complete receiver should be developed as a function of chip rate.

6) Definition and proper weighing of objectives in applying spread spectrum techniques. Possible objectives include
   a. Separation of like coded signals on basis of time of arrival (i.e., to minimize interference or improve capture performance.)
   b. Interference reduction (isolation from signals from unlike coded or narrowband uncoded signals.
   c. Improved performance in multipath environment
   d. Covertness

Weighing of above parameters will effect choice of parameters. For Packet Radio 6.a is likely the most important reason for considering spread spectrum.

7.0 SUMMARY OF DEFINITIONS

$T_b = \frac{1}{\text{Bit Rate}}$ (10 μsec for 100 KBS)

$T_c = \frac{1}{\text{Chip Rate}}$

E_b = Energy per Bit

E_c = Energy per Chip

E_p = Energy of Preamble

N_o = rms noise (bandwidth must be specified)

N - Number of chips per bit (spread factor)

REFERENCES

1) R. L. Kaiser, "Spread Spectrum Considerations", Packet Radio Temporary Note 19 , January 22, 1973

2) R. Gold, "Optimal Binary Sequences for Spread Spectrum Multiplexing", IEEE Transactions on Information Theory, October 1967, pages 619-621

3) M. I. Skolnik, "Introduction to Radar Systems", McGraw-Hill Book Co., 1962, pages 31-35.

4) P. Swerling, "Parameter Estimation for Waveforms in Additive Gaussian Noise", J. Soc. Indust. Appl. Math, Vol. 7, No. 2, June 1959, pgs. 152-166.

5) M. I. Skolnik, "Theoretical Accuracy of Radar Measurements", PGANE Trans., Vol. ANE-7, Dec., 1960, pages 123-129.

6) Bell, Holmes & Ridings, "Application of Surface Wave Technology to Spread Spectrum Communications" This paper will be published on a special Combined Issue of IEEE Transactions on Microwave Theory and Techniques: Sonics and Ultrasonics to be published April, 1973.

R. J. McGuire
April 13, 1973

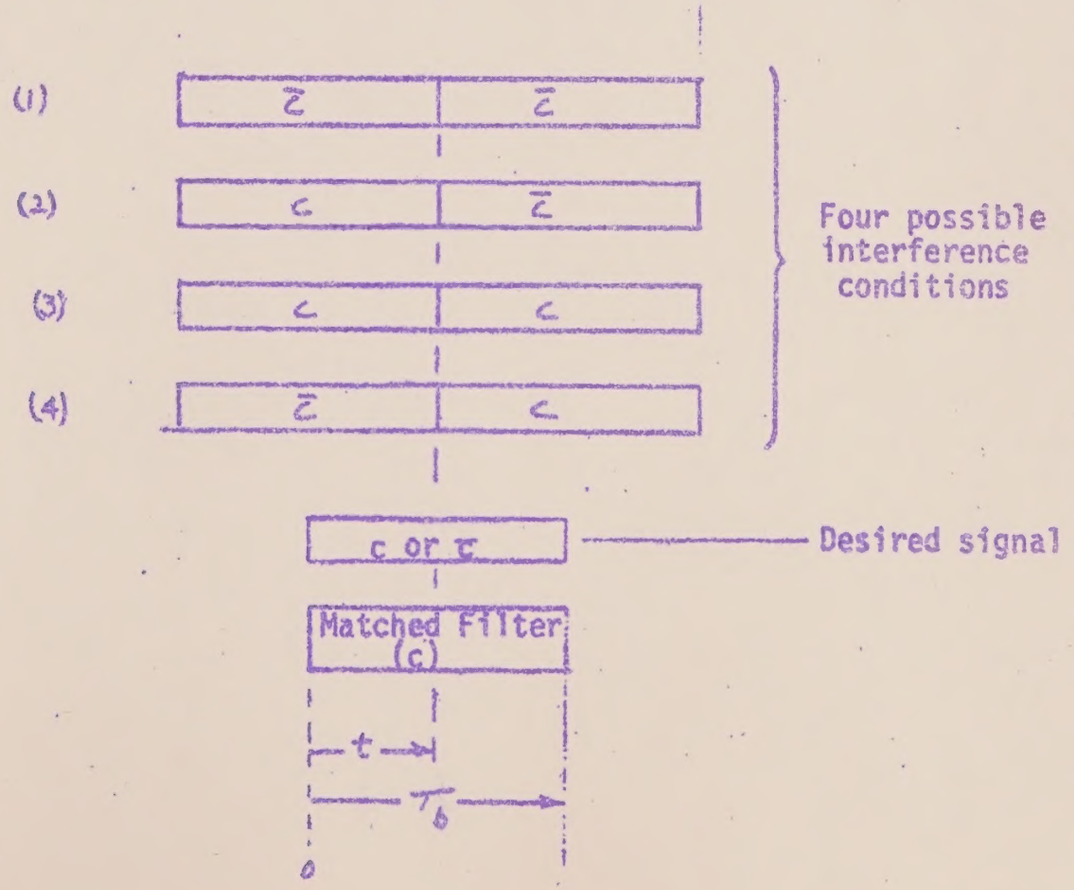
ADDENDUM 1

CORRECTION FOR PACKET RADIO TEMPORARY NOTE #33

1.0 INTERFERENCE LEVELS FOR OVERLAPPING PACKETS

Packet Radio Temporary Note 33 discussed interference levels which could be expected for overlapping packets having the same code assignment. The modulation form which was considered was DPSK. Differential coding was employed which involved transmission of a maximal length code C and its ones compliment $\overline{c}$. The interference level, implied by Figure 2.2-1 for overlapping packets having the same code is incorrect. Interference can be higher, depending on the state of the interferring code (c or $\overline{c}$) and its relative position in time with respect to the desired signal.

To accurately determine the expected interference level introduced by overlapping packet transmissions one must consider the cross-correlation properties of the selected code groups with the matched data filter for the conditions illustrated below.

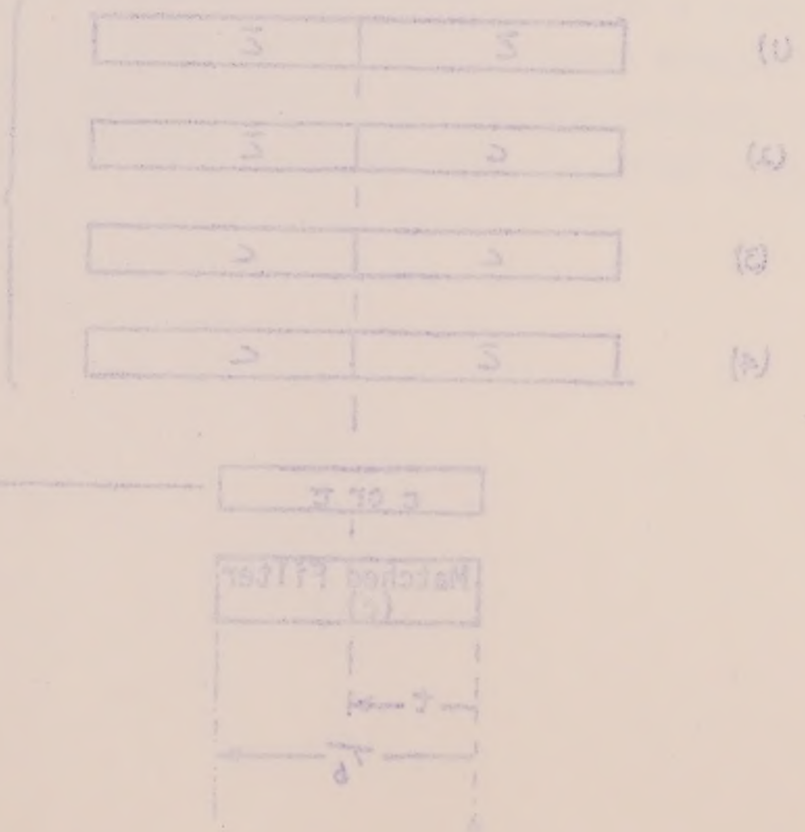


R. J. McGuire
April 13, 1972

ADDENDUM 1

CORRECTION FOR PACKET RADIO TEMPORARY NOTE #33

1.0 INTERFERENCE LEVELS FOR OVERLAPPING PACKETS



Calculations of the interference levels for various codes for values of t where:

$$0 \leq t \leq T_b$$

has not been accomplished as yet. However, it is expected that the rms interferring level will be approximately $\frac{E}{\sqrt{N}}$ for cases 2 and 4 when maximal length codes are employed.

Where: E is the peak value of the auto-correlation function
N is the code length.